LEVEL II

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LOW CARRIER-TO-NOISE RATIO
RECEPTION OF VIDEO DATA LINK
SIGNALS—A LABORATORY INVESTIGATION

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by
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FOREWORD

The work described in this report was performed during fiscal year 1979. This is an interim report of the laboratory investigations conducted on an analog video data link system for operation with low carrier-to-noise ratio signals. The effort was supported by the Naval Air Systems Command and executed under the Strike Warfare Weaponry Technology Block Program under Work Request N0019-79-WR-91001, AIRTASK A03W-03P2/008B/ 9F32-300-000. This AIRTASK provides for continued exploratory development in the air superiority and air-to-surface mission areas.

This report is released at the working level. Because of the continuing nature of this study, it is expected that the results will be extended.

Approved by
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Weapons Department
25 July 1979

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(U) *Low Carrier-to-Noise Ratio Reception of Video Data Link Signals—A Laboratory Investigation*, by W. H. Woodworth. China Lake, Calif., Naval Weapons Center, August 1979. 18 pp. (NWC TP 6129, publication UNCLASSIFIED.)

(U) Techniques involving analog pre-transmission signal processing, wide deviation phase modulation, coherent detection, and field-rate recursive video filtering are combined into a transmission system capable of operating at carrier-to-noise ratios as low as -10 to -17 dB, depending on the type of noise or interference. System design concepts, test results, and planned improvements are described in detail.

CONTENTS

INTRODUCTION

An analog video data link system for operation with low carrier-to-noise ratio signals has been investigated in the laboratory. Based on coherent phase modulation (PM) and video signal processing, the system has great practical value and cost advantages because no complex or expensive components are required at the transmitting end, the expendable part of a tactical missile data link.

This experimental system differs from conventional PM systems in that it employs a simple but effective video signal compaction technique at the transmitter and an adaptive—gain, recursive video filter coupled with a wide-range phase discriminator at the receiver to increase the post-detection signal-to-noise ratio.

In this report we will discuss the theoretical aspects and design details of the system and present data showing satisfactory operation with input carrier-to-noise ratios (C/N_i) down to -10 dB. We will also show how the scheme can be improved so that operation at -20 dB or lower will be possible. Interfacing this system with other modems such as direct sequence pseudonoise and fast frequency hopping will also be discussed.

GENERAL SYSTEM DESCRIPTION

Analog RF transmission employing amplitude modulation (AM) or PM can be coherently demodulated even though C/N_i is very low, because the carrier can be recovered with a very narrow bandwidth filter or phase-locked loop arrangement and then multiplied by the incoming signal to recover the modulation. In this scheme, the output signal-to-noise ratio (S/N_o) cannot exceed C/N_i as opposed to frequency-modulated (FM) systems where improvement is obtained when signals are above the C/N_i threshold, typically about +9 dB or so. However, with FM, S/N_o falls rapidly when operating below the C/N_i threshold, becoming substantially lower than AM or PM under the same conditions.[1]

In recent years, systems have been proposed and constructed by a number of different groups using PM for video transmission with the intention of operating at low C/N_i.[2,*] However, even though the PM receiver continues to coherently demodulate, the visual display becomes generally unusable below 0 dB C/N_i. Therefore, in an attempt to operate substantially below 0 dB, we have devised a pre-modulation video processing method at the transmitter which increases the effective

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* Informal discussions with Hughes Aircraft Co. personnel.

[1] R. F. Panter. *Modulation, Noise and Spectral Analysis.* New York, McGraw-Hill, 1965.

[2] Resdel Engineering Co. Study Contract to NWC, 1972, *Phase Modulated Video System.*

signal modulation more than 3 dB by eliminating low spatial frequency vertical information and reducing the video bandwidth to 1.75 MHz with horizontal edge enhancement. At the receiving end we have increased the usable phase deviation range and added a digital recursive video filter to raise the post-detection S/N_o significantly. In this later device, the video signal is recirculated through a one-field or one-frame delay line so that the field-to-field or frame-to-frame coherent image components are reinforced while random or nonsynchronous components are suppressed. Interestingly, the amount of recursive gain in the filter can be made a function of C/N_i, a distinct advantage because the recursive filter adds "smear" to the picture which increases with increasing gain. Thus under high C/N_i conditions, the picture will have no smear because the recursive gain will be low and as C/N_i deteriorates, the recursive gain can be automatically raised, creating smear but maintaining high S/N_o. It should be noted that the adaptive gain feature was not included in this effort, but the design has been completed and appears to be feasible in all respects.

Another aspect of the system which we have not yet investigated is the problem of synchronizing the digital recursive filter (and also the video). However, it appears that several methods are feasible based on either wideband or narrowband synchronizing signal encoding.

VIDEO SIGNAL PROCESSING

We have found that for most airborne video applications, a reduction in the amount of displayed video information is not only tolerable, but highly desirable, particularly when the display size is small. The point to understand is that a normal, full-bandwidth video system provides the viewer with more information than can be assimilated without detailed examination of the picture in a calm environment. Consequently, we reduced the video bandwidth to 1.75 MHz (from the normal 5 MHz), eliminating small scene details, with the "fuzzy" edges resulting from the restriction being restored by an edge-enhancement process which does not increase the video bandwidth. Visually, most observers prefer the lower bandwidth picture because it appears "sharper" and the larger scene details are more visible. A simplified diagram of the edge enhancement process is shown in Figure 1.

It should be noted that, to avoid bandwidth shrinkage, the linear phase filter shown in Figure 1 must actually be distributed throughout the entire system rather than being lumped in one location. The most effective arrangement places underdamped response poles at the transmitting end (a form of pre-emphasis), with the receiver bandpass and post-detection filtering locating the remaining poles.

Low-spatial frequency vertical information also can be eliminated without visual loss because actually, very little large-area vertical information is useful to a viewer of air-to-ground imagery. We all know the sky is at the top and when viewing with less oblique angles, most scenes are quite uniform, top to bottom. Figure 2 shows the vertical background removal scheme.

Here, 512-bit bucket-brigade analog delay line delays the average value of each video line by 512 lines, and the 500-Hz input filter adds another 13 lines giving 525 lines total delay, or one full frame (1/30 second). This low-frequency vertical background signal is subtracted from the real-time video, suppressing the low-spatial frequency vertical picture components.

The result of the horizontal and vertical processing is to reduce the peak-to-peak video amplitude by 3 to 6 dB or more depending on the scene characteristics. An automatic gain control (AGC) system raises the level of useful components and, in effect, increases the modulation level of transmitted signal, which is a direct S/N increase at the receiver output.
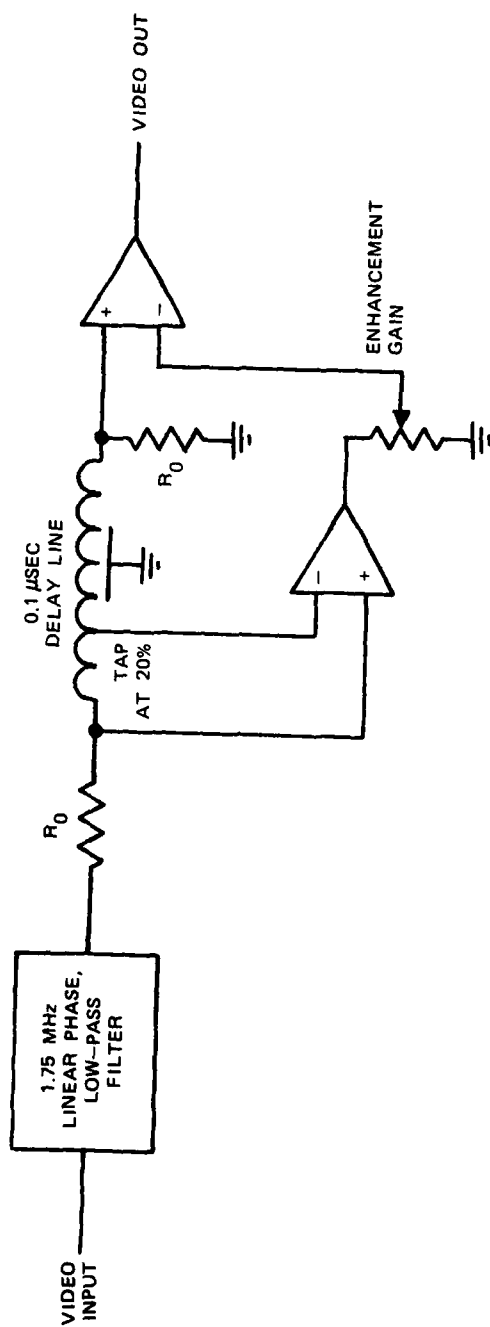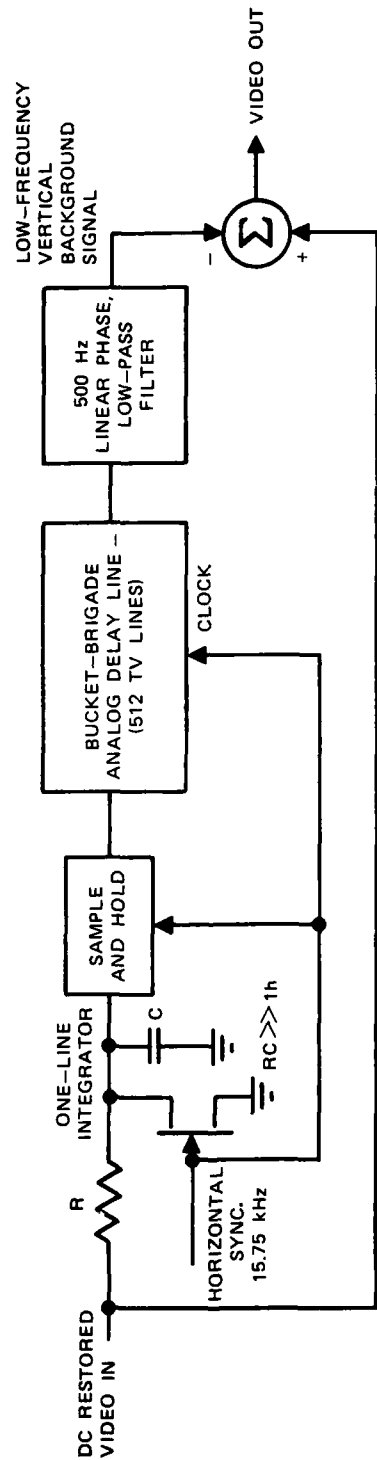
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FIGURE 1. Block Diagram, Horizontal Edge Enhancement.

FIGURE 2. Low-Frequency Vertical Background Removal System.

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THE PHASE MODULATION SYSTEM

A necessary part of this video transmission system is the phase modulation transmitter and receiver. For simplicity we operate at a frequency of 60 MHz, which could be the intermediate frequency of a microwave receiver, avoiding the complexities of additional mixers, etc. Phase modulation is accomplished by driving an active, double-balanced mixer with the pre-processed video, and then summing with a quadrature carrier followed by a limiting amplifier driving the output line to the receiver. A simplified diagram of the modulator is shown in Figure 3.

Another type of video phase modulator that can be used with existing FM transmitters has also been devised and tested wherein we phase modulate by frequency-modulating with the derivative of the signal, as shown in Figure 4.

The parallel LRC network performs the input differentiation and generates a pair of complex response poles, normally underdamped, as part of the overall system response. The high-frequency response is given by:

$$F(s) = \frac{s\frac{L}{R_1}}{s^2 LC_2 + s\frac{L}{R_1} + 1}$$

The L/R time constant of the inductor causes a rise in the low-frequency phase modulation response and is compensated for by the input network C_1 and R_1. In this case $L/R_2 = R_1 C_1$ for flat response. As mentioned before, this modulation scheme can be used with existing FM video transmitters if the residual phase noise or incidental FM is low, a situation normally prevailing with crystal-controlled FM transmitters.

Two different receiver designs have been tested. The first uses a conventional phase locked loop of 1-KHz bandwidth to demodulate the signal, as shown in Figure 5.

The second receiver design is less conventional and uses a crystal filter for carrier recovery, combined with an automatic frequency control (AFC) loop for carrier tracking, as shown in Figure 6. The AFC loop keeps the IF frequency within the 400-Hz crystal filter bandpass, and demodulated video is taken from the phase discriminator output at the point where the AFC error signal is taken. High-frequency bandwidth is limited by the IF bandpass and post-detection filtering, and the low-frequency bandpass is determined by the AFC closed-loop response. As mechanized here, the loop bandwidth is about 150 Hz, and the calculated low-frequency response shows a slight peak at this frequency followed by a 6-dB/octave roll off below. It should be noted that the dynamic range of active components and phase discriminators must be sufficient to accommodate the expected signal level plus in-band noise or interference; otherwise, unwanted signal suppression will occur due to limiting action. In the AFC loop, however, a hard-limiter is deliberately placed at the output of the crystal filter so that the error signal output from the frequency discriminator becomes a linear function of signal amplitude rather than the square function without limiting. A second phase discriminator preceded by a hard-limiter has also been included. Here, in-band noise will cause signal suppression at the output, and since known output signal component amplitudes (i.e., synchronizing signals) can be measured, we obtain a direct measure of S/N_0 which can be used to control the recursive filter gain. No further design effort was expended on the receiver since both types appeared to be functioning properly.
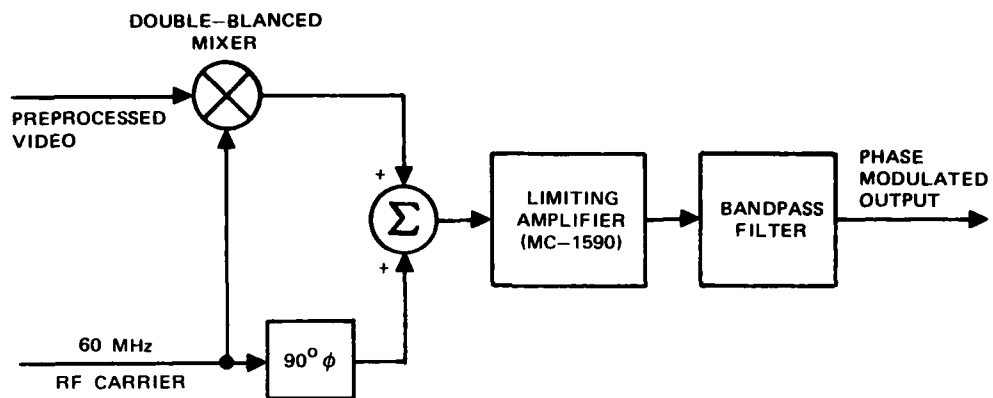
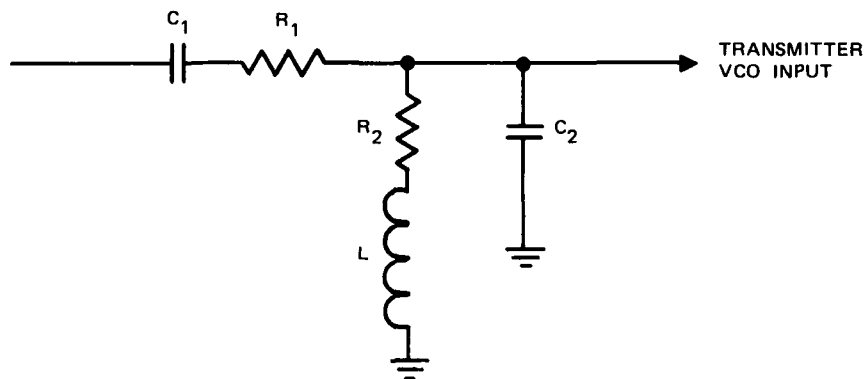FIGURE 3. Phase Modulation Using Double-Balanced Mixer.
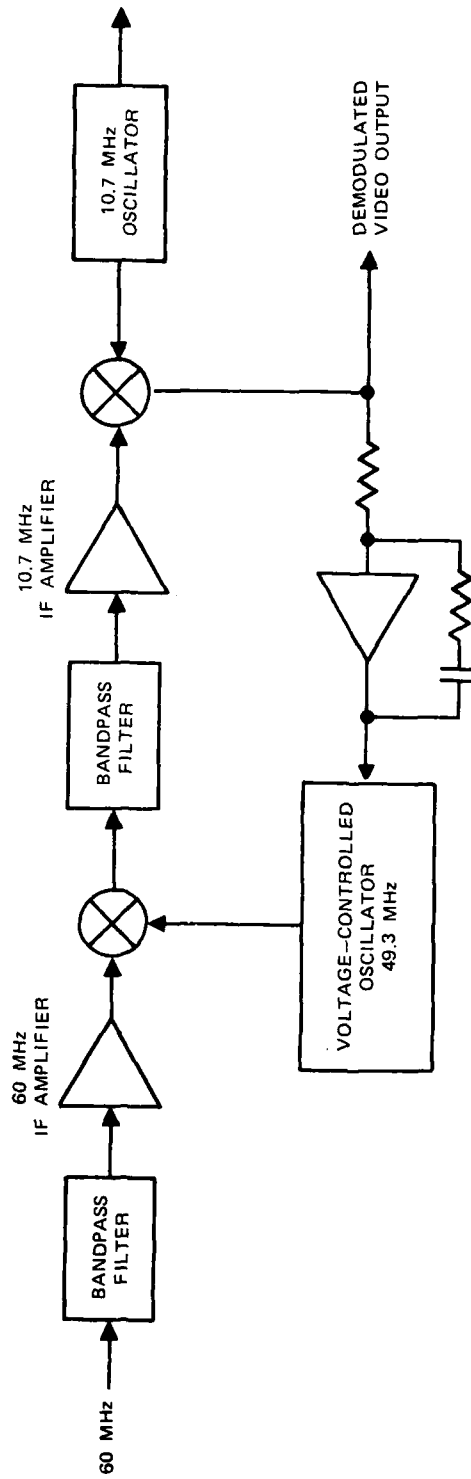

FIGURE 4. Phase-Modulating an FM Transmitter.

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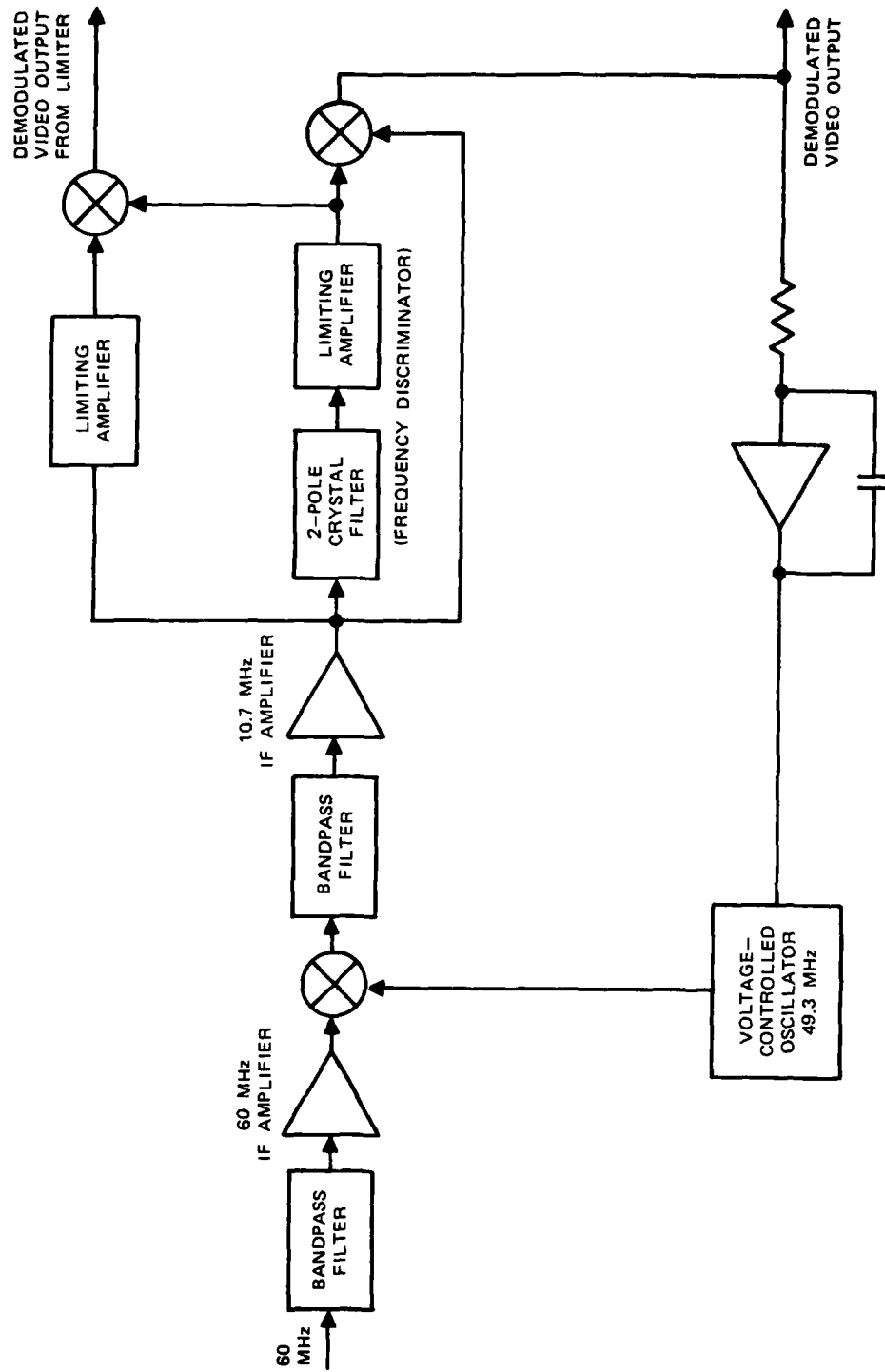FIGURE 5. Phase-Locked Loop Demodulator.

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FIGURE 6. Crystal Filter Demodulator With Automatic Frequency Control.

A significant improvement to S/N_o can be made by employing a novel phase demodulation scheme in the receiver. Whereas conventional demodulators are essentially linear up to ±1 radian deviation (B = 1) and limit at ±90 degrees, this new demodulation scheme operates linearly to ±2 radians (B = 2), limiting at ±180 degrees. We have devised two different demodulation methods for the higher B signal and both appear to be practical. In Figure 7a, at the receiver, an in-phase carrier is added to the PM signal reducing the signal phase deviation to less than ±90 degrees. If ϕ_s is limited to ±2 radians deviation and ϕ_m to ±90 degrees, then ρ can have a maximum value of 2.4. A minimum value of $\rho = 0.416$ occurs where the rate-of-change of ϕ_m goes to zero as ϕ_s approaches ±2 radians. These are wide limits and should present no signal level tolerance problems. Following carrier addition, the signal is demodulated as conventional PM.

A second method is shown in Figure 7b, where frequency division is used to reduce B by a factor of 2, the maximum possible amount. Here the bandpass filter can be the phase locked loop or passive crystal filter of the previously described receivers.

The S/N_o improvement obtained with these demodulators can be calculated from the ratios of power in the first pair of sidebands for the different values of B. In the case of B = 1, the first sideband power is 40% of the unmodulated carrier; and with B = 2, 70%, giving an improvement ratio of 1.75 or 2.4 dB. However, with wide deviation, increasing the bandpass to cover the next pair of sidebands does not increase S/N_o even though more demodulated output is obtained because the input noise power increases faster than the sideband power; hence, increasing bandwidth is not effective. However, for modulating frequencies of one-half maximum and less, more sidebands are within the bandpass and the S/N_o increase for these can approach 6 dB.

It shoud be noted that with B = 2, the carrier power is reduced to 5% of the unmodulated carrier compared to 60% with B = 1, or a factor of 12. Thus, in order to preserve the same S/N_o from the carrier filter, the carrier recovery bandpass must be reduced by this factor. For the receivers previously described, the bandpass is already sufficiently narrow to accommodate the reduction when C/N_i is -20dB.
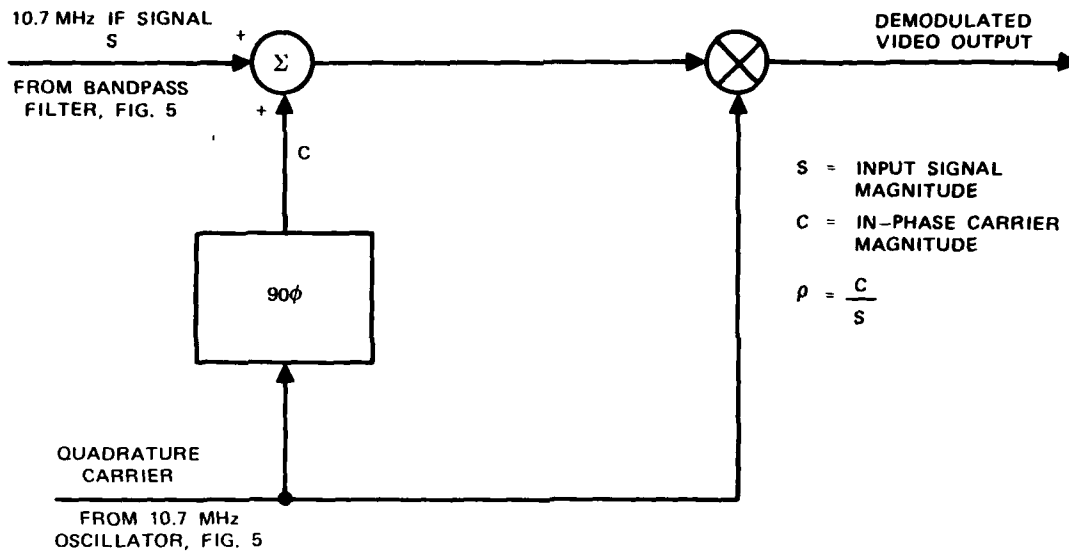
THE RECURSIVE VIDEO FILTER

Recursive filtering is a generic name given to a class of filters that have positive output-to-input feedback around a delay line giving response peaks at frequencies n/d where n is an integer 1,2,3. . . . and d is the line delay in seconds. Thus, an input signal with frequency components n/d will be passed by the filter while random or nonsynchronous signals will be attenuated. Figure 8 shows the filter in block diagram form.
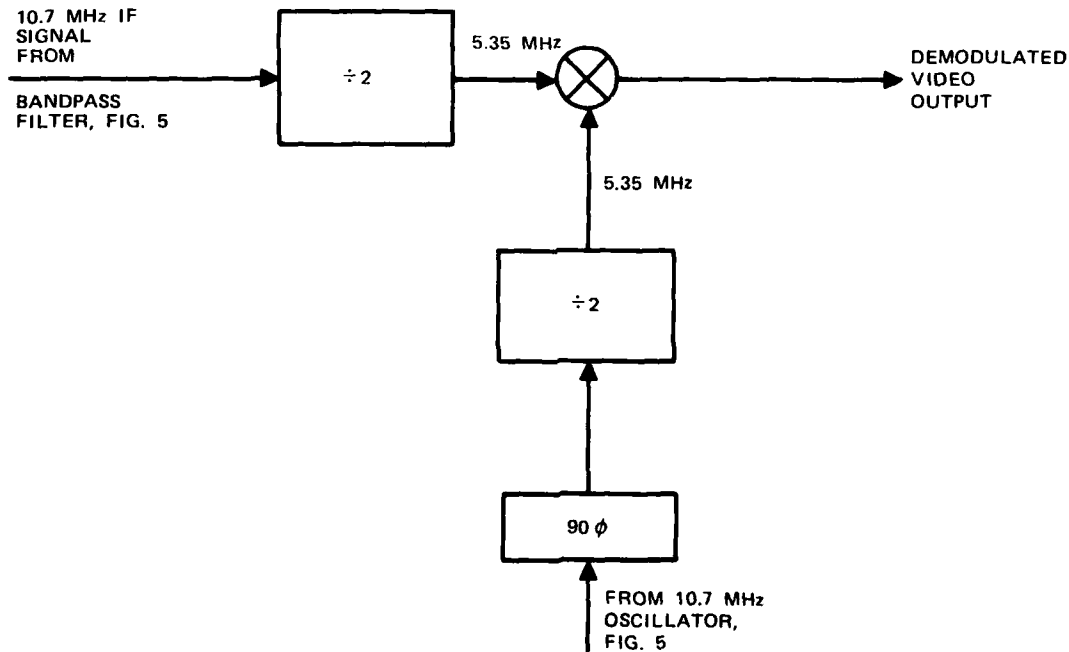
It is not difficult to show that the S/N_o improvement obtained for matching signals is:

$$I = \sqrt{2a - 1}$$

where "a" is gain parameter of Figure 8. Another way of looking at the filter is to consider the output to be the time average of N periodic signals so that the S/N_o improvement is $\sqrt{N}$ and as a

10.7 MHz IF SIGNAL
S

FROM BANDPASS
FILTER, FIG. 5

+
+

Σ

C

90ϕ

QUADRATURE
CARRIER

FROM 10.7 MHz
OSCILLATOR, FIG. 5

$\otimes$

DEMODULATED
VIDEO OUTPUT

S = INPUT SIGNAL
MAGNITUDE

C = IN-PHASE CARRIER
MAGNITUDE

$\rho = \dfrac{C}{S}$

(a) In-phase carrier addition.

10.7 MHz IF
SIGNAL
FROM

BANDPASS
FILTER, FIG. 5

$\div 2$

5.35 MHz

$\otimes$

DEMODULATED
VIDEO
OUTPUT

5.35 MHz

$\div 2$

90 ϕ

FROM 10.7 MHz
OSCILLATOR,
FIG. 5

(b) By carrier frequency division.

FIGURE 7.   Reduced Phase Deviation.

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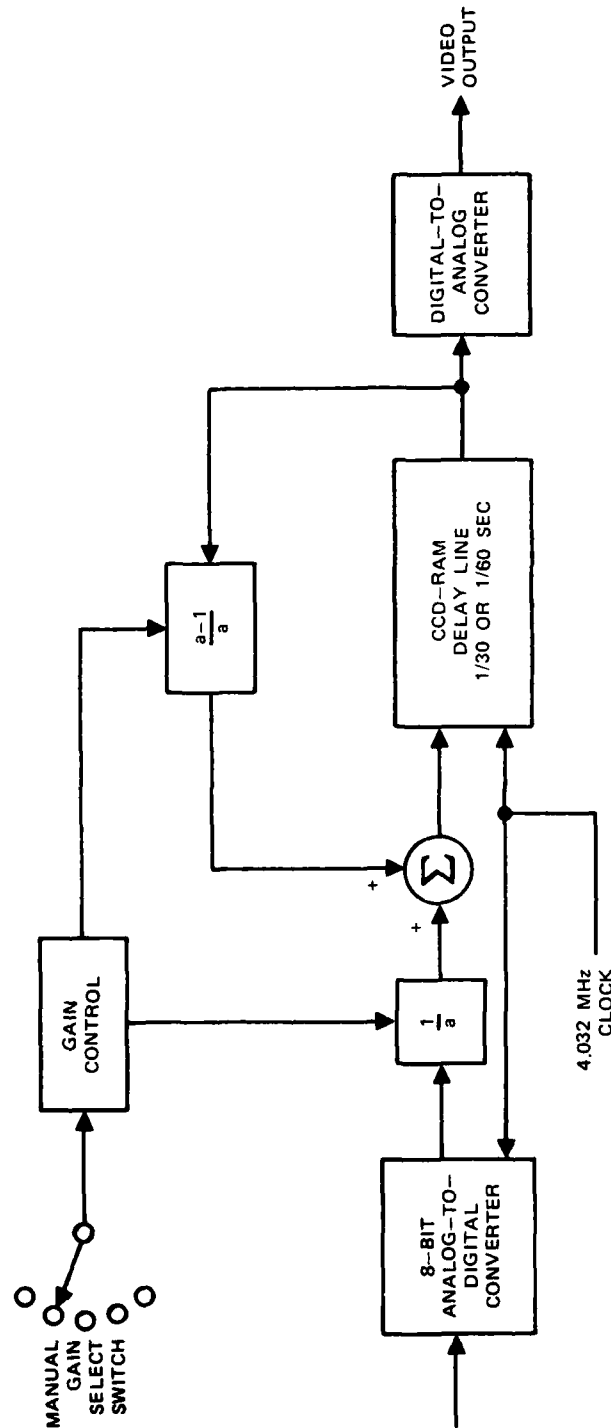
FIGURE 8. Block Diagram Recursive Video Filter.

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SYNCHRONIZATION

The recursive filter delay must be held very close to one field or frame period in order to avoid resolution loss and poor noise suppression due to lack of correlation between input and output. In the digital delay line used here, the time delay is determined strictly by the clock frequency and the number of elements in the line; hence, it can be adjusted by varying the clock frequency. Furthermore, the timing can be made exact if the required frequency clock is an exact multiple of the video horizontal line rate. For the filter here, the clock frequency is 4.032 MHz, 256 times the horizontal line rate of 15.750 KHz. Because of this precise relationship, a number of possible schemes for deriving the clock frequency from the transmitted horizontal synchronizing signal can be envisioned. It should be noted that we need no vertical synchronizing signal for the filter because the delay line is exactly 525 video lines long (one frame) or alternately 262 and 263 for a video field delay.

An obvious timing method is to transmit a binary code in the video horizontal blanking interval, replacing the normal synchronizing pulse. For example, a 32-bit code word will occupy 8×10^{-6} seconds at a 2.016-MHz bandwidth (4.032-MHz bit rate), a value easily fitted into a nominal 12×10^{-6}-second blanking interval. Correlation with a matched filter or similar device at the receiving end increases the S/N_0 by $\sqrt{32}$ or 15 dB and another 5 dB can be obtained from a narrow bandwidth filter following the correlator output, thus raising the processing gain to 20 dB to match the intended video gain. Another method of increasing S/N_0 would be to use a separate line-rate recursive filter before or after correlation followed by a phase locked loop tracking the output correlation pulse and from it, synthesizing the 4.032-MHz clock frequency.

Another scheme uses the output from the video recursive filter set at high gain to obtain the necessary suppression. However, the accuracy of the free-running clock required to maintain high recursive gain becomes impractical; much better than 1 part per million (PPM) is needed.

As a final consideration, if some method of PN bi-phase or fast-frequency-hopping is adapted, the clock frequency could be extracted from the decoding system which, in turn, could be related to the IF carrier frequency, thus tying the entire system timing together. Of course, the transmission is automatically encrypted in this case and the carrier is given protection from interferrence by the processing gain of the particular modem selected.

RESULTS OF LABORATORY MEASUREMENTS

In order to determine the performance of the phase modulated system with recursive video filtering, an experimental system was set up in the laboratory using the crystal filter demodulator. As depicted in Figure 9, video from a camera is the input to the phase modulator, the double-balance mixer plus quadrature carrier-type in this case. The IF noise source is a 60-MHz voltage-controlled oscillator driven by a wideband noise generator. After bandpass limiting, the summed signal-plus-noise is either measured by an RF power meter or directed to the demodulator input. Measurements were made with C/N_i very high, 0 dB and −10 dB with the results shown in Figures 10a through 10f. It is clear that the recursive filter improves the visual output signal-to-noise ratio by at least 10 dB with random noise as the interfering signal. We realize that presenting valid data in the form of
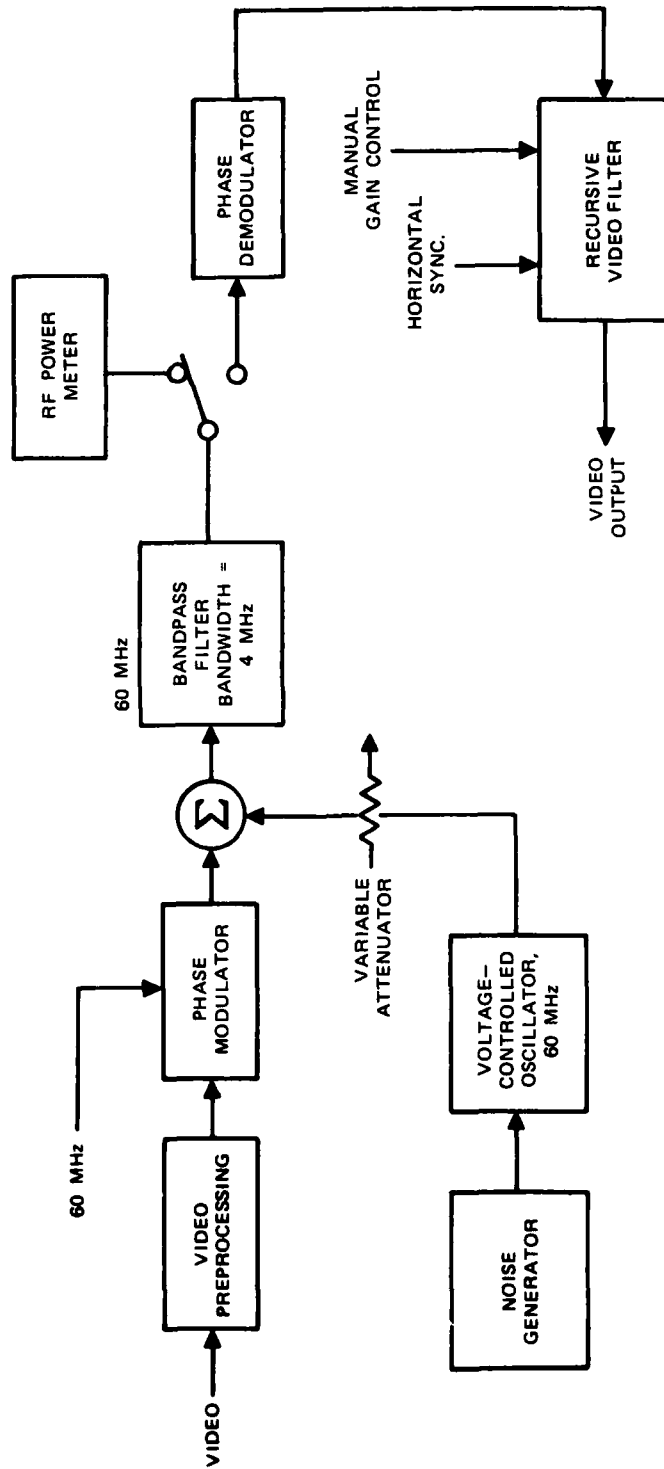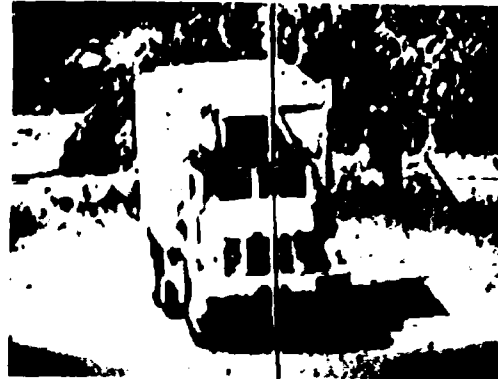
FIGURE 9. Experimental Test Setup.
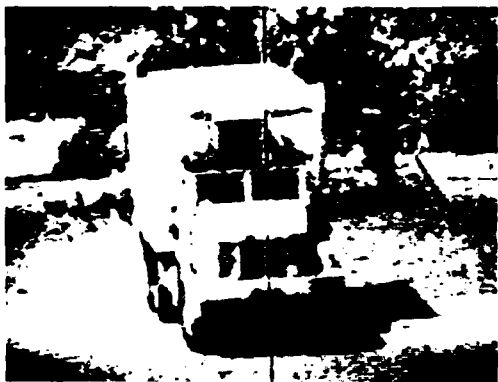
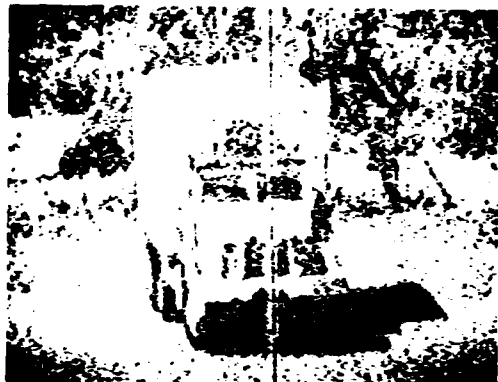WITH RECURSIVE FILTER   WITHOUT RECURSIVE FILTER
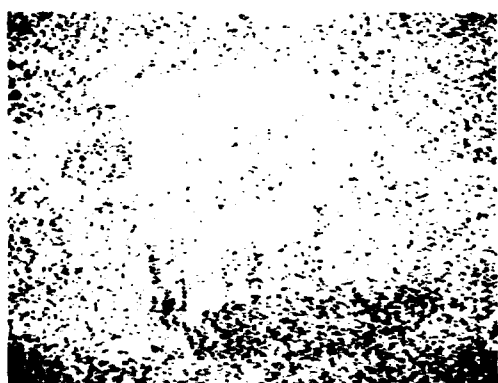

(a) C/N$_i$ = +30 dB

(b) C/N$_i$ = +30 dB

(c) C/N$_i$ = 0 dB

(d) C/N$_i$ = 0 dB

(e) C/N$_i$ = -10 dB

(f) C/N$_i$ = -10 dB

FIGURE 10.   Phase-Modulated Video Display, Photographs
With Decreasing Carrier-to-Noise Ratios (C/N$_i$).

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photographs of the display monitor requires the exposure time to be approximately one frame (1/30 second), and overpressure or underexposure must be avoided if the gray scale is to be preserved. We believe that the photographs of Figure 10 very closely approximate the true visual appearance of the display in all aspects except for image smear with motion—hard to show with stills.

Testing with interference other than random noise was also carried out. In the case of in-band CW, the visual improvement with the recursive filter was at least as good as that obtained for random noise and possibly 3 dB better. Also, short duty cycle IF pulses (10% and less) are very effectively processed out by the phase demodulator and filter, a situation quite different from FM systems where high-amplitude pulses capture the FM demodulator, causing high-amplitude video output during the pulse duration. With phase modulation, the demodulated pulse appears as a "beat note" between the pulse and carrier frequencies with zero average value at the demodulator output. By adding the recursive filter we can operate satisfactorily with C/N_i at -17 dB, based on the average carrier power (not peak), and with very high pulse power, IF limiting improves the performance even more.

SUMMARY AND CONCLUSIONS

In this report, we have disclosed a video transmission system with these important characteristics:

1. Demonstrated capability to transmit usable video with carrier-to-interference ratios (C/N_i) less than -10 dB for random noise, -13 dB for CW, and -17 dB for pulses (based on average power).

2. Use of video pre-processing, RF phase modulation, and coherent detection followed by recursive video filtering to raise S/N_o.

3. Inherently low cost. Relatively simple signal processing and RF components at the transmitter and straightforward digital processing at the receiver keep the overall complexity low compared to systems using video transforms and wideband RF modems.

4. Capability for improved operation with a 12-bit digital memory and more complex modems.

Based on the results of laboratory testing, it can be concluded that a video transmission system capable of operating at low carrier-to-noise ratios can be designed so that only relatively simple components are required at the transmitting end and with modest complexity at the receiving end. Synchronization of the recursive filter was a problem not addressed experimentally, but two possible methods have been identified with high probabilities of being successful; hence, we feel that solving this problem should be the primary objective of the next phase of development. With the facilities available, flight testing of a complete system will be relatively straightforward, placing emphasis on performance in a noisy environment and the effects of multipath propagation on the carrier recovery and synchronization subsystems.